A SLIDING MODE CONTROL FOR POSITIVE OUTPUT ELEMENTARY LUO CONVERTER

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Abstract: This paper studies a design of Sliding Mode Control (SMC) for Positive Output Elementary Luo Converter (POELC) operated in Continuous Conduction Mode (CCM). The POELC performs the voltage conversion from positive DC source voltage to positive DC load voltage. Due to the time varying and switching nature of POELC, its dynamic behavior becomes highly non-linear. In order to improve dynamic performances of POELC for both static and dynamic specifications, SMC is proposed. The SMC is designed by using state-space average modeling of POELC. The simulation of POELC with its control model is implemented in MatLab/Simulink. The performances of proposed controller with system are tested at various regions. The simulation results validate the effect of SMC on the static and dynamic performances of POELC.

Key words: DC-DC converter, positive output elementary luu converter, state-space average method, sliding mode control.

1. Introduction

DC-DC conversion technology has been developing very rapidly, and DC-DC converters have been widely used in industrial applications such as dc motor drives, computer systems and communication equipments.

Switch-mode power converters represent a particular class of Variable Structure Systems (VSS), and they can take advantage of nonlinear control techniques developed for this class of system [1]. An ideal control should ensure system stability in any operating condition and good static and dynamic performances in terms of rejection of input voltage disturbances and load changes [2].

These characteristics, of course, should be maintained in spite of large input voltage, output current, and even parameter variations (robustness) [3].

A classical control approach relies on the state-space averaging method, which derives an equivalent model by circuit-averaging all the system variables in a switching period [4-6]. From the average model, a suitable small signal model is then derived by perturbation and linearization around a precise operating point. Finally, the small-signal model is used to derive all the necessary converter transfer functions to design a linear control system by using classical control techniques [7]. The design procedure is well known, but it is generally not easy to account for the wide variation of system parameters, because of the strong dependence of small-signal model parameters on the converter operating point. Multi-loop control techniques, such as current-mode control, have greatly improved power converter dynamic behavior, but the control design remains difficult especially for high-order topologies.

Sliding Mode Control (SMC), which is derived from variable structure system theory, extends the properties of hysteresis control to multivariable environments, resulting in stability even for large supply and load variations, good dynamic response, and simple implementation over the conventional controller. A number of contributions to SMC of power converters with diverse sliding surfaces are available [8-10].

The POELC is a new series of DC-DC converters possessing high-voltage transfer gain, high power density; high efficiency, reduced ripple voltage and current [11]. The voltage lift technique has been successfully employed in the design of DC/DC converters, e.g. three-series luo-converters, in which the output voltage increases stage-by-stage in arithmetic progression [11-12]. However, their circuits are complex. An approach, positive output elementary luo converters, that implements the output voltage increasing in arithmetic progression with a simple structured have been introduced. These converters also effectively enhance the voltage transfer gain. These converters are widely used in computer peripheral equipment, switch mode power supply, medical equipments and also industrial applications, especially for high-voltage projects [11]. Many literatures have reported the general design issues of SMC in dc-dc converters like buck, cuk, and buck–boost converters.
[13-14]. Intensive research in the area of DC-DC converter has resulted in novel circuit topologies and these converters in general have complex non-linear models with parameter variation in [15].

Direct regulation/tracking control of the output voltage for POELC results in a non-minimum phase system and therefore an unstable controller. An attempt is made to show that controlling the current can indirectly control the output voltage in converter.

Therefore in this paper, the voltage regulation for POELC operated in CCM using SMC is proposed. The state-space average model for POELC is derived at first and SMC is designed. The detailed discussion of hitting, existence and stability conditions of the SMC for POELC is studied in this paper. The performances of the controller in terms of robustness and dynamic response will be improved by proper selection of the controller gains. Section 2, present the operation and mathematical model of POELC. The design of SMC for POELC is presented in section 3. The design computation of POELC circuit components and the controller gains is well executed in section 4. Simulation results of system at various regions are discussed in section 5. The conclusions and future work of system is discussed in section 6

2. Operations and Mathematical Modeling of POELC
2.1 Operation of POELC

The POELC circuit is shown in Fig. 1. It consists of input dc supply voltage $V_{in}$, power switch S (N-channel MOSFET), positive Luo pump S-$L_1$-$C_1$-$D$ and low pass filter $L_2$-$C_2$, d duty cycle, load resistance R, $V_o$ average output voltage and $I_o$ average output current.

Fig. 1. Circuit diagram of POELC.

In the description of the converter operation, we assume that all the components are ideal and that the POELC operates in a CCM. Figs. 2 and 3 show the two topological modes for a period of POELC operation [11-12].

In mode 1 operation, when the switch is ON, the inductor $L_1$ is charged by the supply voltage $V_{in}$. At the same time, the inductor $L_2$ absorbs the energy from source and the capacitor $C_1$. The load is supplied by the capacitor $C_2$. The equivalent circuit of POELC in mode 1 operation is shown in Fig. 2.

Fig. 2. Equivalent circuit of POELC in Mode 1 operation.

During mode 2 operation, switch is in OFF state, and hence, the current $i_o$ drawn from the source becomes zero, as shown in Fig. 3. Current $i_{L1}$ flows through the freewheeling diode D to charge capacitor $C_1$. Inductor $L_1$ transfers its stored energy to capacitor $C_1$. Current $i_{L2}$ flows through the capacitor $C_2$ and load resistance R, freewheeling diode D to keep itself continuous and both currents $i_{L1}$ and $i_{L2}$ decrease.

Fig. 3. Equivalent circuit of POELC in Mode 2 operation.

Therefore, the voltage transfer gain of POELC in continuous conduction mode is given by equation (1)

$$G = \frac{V_o}{V_{in}} = \frac{d}{1-d}$$

and the average output current is

$$i_o = \frac{1-d}{d}i_{in}$$
2.2 State Space Averaging Model

The state–space modeling of the equivalent circuit of the positive output elementary Luo converter with state variables \( i_{L1}, \ V_{C1}, \ i_{L2} \) and \( V_{C2} \) as follows [16-17]

\[
\begin{bmatrix}
\frac{di}{dt} \\
\frac{dv}{dt}
\end{bmatrix} =
\begin{bmatrix}
0 & \frac{1}{L} & 0 & 0 \\
-\frac{1}{C} & 0 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
i_{L1} \\
v_{C1}
\end{bmatrix} +
\begin{bmatrix}
\frac{-V_{C1} + V_{in}}{L} \\
\frac{i_{L1} + i_{L2}}{C}
\end{bmatrix}
\gamma +
\begin{bmatrix}
0 \\
0
\end{bmatrix}
\]

\[\text{V}_{o} = v_{4}\]

\[v = Av + B\gamma + C\]  

Where \( \gamma \) is the status of the switches, \( v \) and \( b \) are the vectors of the state variables (\( i_{L1}, \ V_{C1}, \ i_{L2} \) \( V_{C2} \)) and their derivatives respectively,

\[
\gamma = \begin{cases} 
1 \rightarrow S \rightarrow \text{ON} \\
0 \rightarrow S \rightarrow \text{OFF}.
\end{cases}
\]

3. Design of Sliding Mode Controller

In sliding mode theory, the SMC requires sensing of all state variables of POELC and generation of suitable references for each of them [19]. According to principle of the SMC, capacitors voltages \( V_{C1} \) and \( V_{C2} \) are made to follow as faithfully as possible as to its references. However, the inductor current reference is difficult to evaluate since that generally depends on load power demand supply voltage, and load voltage. To overcome this problem, in implementation the state variable errors for the inductor current (\( i_{L1} - i_{L1, ref} \)) and (\( i_{L2} - i_{L2, ref} \)) can be obtained from feedback variable \( i_{L1} \) and \( i_{L2} \) by means of a high-pass filter in the assumption that their low-frequency component is automatically adapted to actual converter operation. Thus, only the high-frequency component of this variable is needed for the control. This high pass filter increases the system order and can heavily alter the converter dynamics. In order to avoid this problem, the cut-off frequency of the high-pass filter must be suitably lower than the switching frequency to pass the ripple at the switching frequency, but high enough to allow a fast converter response [19].

In the design of the POELC operated in CCM, the following are assumed:

- ideal power switches
- power supply free of dc ripple
- converter operating at high-switching frequency

To require a good output voltage regulation of POELC, the sliding surface equation in the state space, which is expressed by a linear combination of state-variable errors (\( \varepsilon_{1}, \varepsilon_{2}, \varepsilon_{3} \) and \( \varepsilon_{4} \)) (respective differences of feedback reference current/voltage and feedback current/voltage), must be selected optimally [18-19].

\[
S(i_{L1}, V_{C1}, i_{L2}, V_{C2}) = K_{1}\varepsilon_{1} + K_{2}\varepsilon_{2} + K_{3}\varepsilon_{3} + K_{4}\varepsilon_{4}
\]

Where coefficients \( K_{1}, \ K_{2}, \ K_{3} \) and \( K_{4} \) are proper gains, \( \varepsilon_{1} \) is the feedback current error, \( \varepsilon_{2} \) is the feedback voltage error, \( \varepsilon_{3} \) is feedback current error and \( \varepsilon_{4} \) is the feedback voltage error, or

\[
\varepsilon_{1} = i_{L1} - i_{L1, ref}
\]

\[
\varepsilon_{2} = V_{C1} - V_{C1, ref}
\]

\[
\varepsilon_{3} = i_{L2} - i_{L2, ref}
\]

\[
\varepsilon_{4} = V_{C2} - V_{C2, ref}
\]

\[
S(i_{L1}, V_{C1}, i_{L2}, V_{C2}) = K_{1}(i_{L1} - i_{L1, ref}) + K_{2}(V_{C1} - V_{C1, ref})
\]

\[
+ K_{3}(i_{L2} - i_{L2, ref}) + K_{4}(V_{C2} - V_{C2, ref})
\]

Fig. 4. Principle scheme of SMC applied to POELC.

The signal \( S (i_{L1}, V_{C1}, i_{L2}, V_{C2}) \) generated using equation (11) while applied to conventional hysteresis modulator generates the gate pulses to MOSFET switch. The resulted control arrangement is shown in Fig. 4. Status of the switch \( \gamma \) is controlled by hysteresis
block H, which aims to minimize the error of variables $i_{L1}$, $V_{C1}$, $i_{L2}$, and $V_{C2}$.

The system response is determined by the circuit parameters and coefficients $K_1$, $K_2$, $K_3$ and $K_4$. With a proper selection of these coefficients in any operating condition, high control robustness, stability, and fast response can be achieved.

### 3.1 Selection of Control Parameters

Once the POELC parameters are selected, inductances $L_1$ and $L_2$ and are designed from specified input and output current ripples, capacitors $C_1$ and $C_2$ are designed so as to limit the output voltage ripple in the case of fast and large load variations, and maximum switching frequency is selected from the proposed converter ratings and switch type. The system behavior is completely determined by coefficients $K_1$, $K_2$, $K_3$ and $K_4$, which must be selected so as to satisfy existence and ensure stability and fast response, even for large supply variations, load variations, set point variations and also components variations.

According to the variable structure system theory, the converter equations must be written in the following form [18-19]:

$$x = Ax + B\gamma + D$$  \hspace{1cm} \text{(12)}

Where $x$ represents the vector of state-variables errors, given by

$$x = V - V^*$$  \hspace{1cm} \text{(13)}

Where $V^* = \begin{bmatrix} i_{L1ref} & V_{C1ref} & i_{L2ref} & V_{C2ref} \end{bmatrix}^T$ is the vector of references.

By substituting equation (3) in equation (14), one obtains

$$D = AV^* + C$$  \hspace{1cm} \text{(14)}

$$D = \begin{bmatrix} 0 & \frac{1}{L_1} & 0 & 0 \\ -\frac{1}{C_1} & 0 & 0 & 0 \\ 0 & 0 & \frac{1}{L_2} & 0 \\ 0 & \frac{1}{C_2} & 0 & -\frac{1}{RC_2} \end{bmatrix}$$

Substituting equation (13) in equation (11), the sliding function can be rewritten in the form

$$S(x) = K_1x_1 + K_2x_2 + K_3x_3 + K_4x_4 = K^T x$$  \hspace{1cm} \text{(16)}

Where $K^T = [K_1, K_2, K_3, K_4]$ and $x = [x_1, x_2, x_3, x_4]^T$.

The existence condition of the sliding mode requires that all state trajectories near the surface be directed toward the sliding plane. The controller can enforce the system state to remain near the sliding plane by proper operation of the converter switch.

To make the system state move toward the switching surface, it is necessary and sufficient [19]

$$\begin{cases} S(x) < 0, \text{ if } S(x) > 0 \\ S(x) > 0, \text{ if } S(x) < 0 \end{cases}$$  \hspace{1cm} \text{(17)}

SMC is obtained by means of the following feedback control strategy, which relates to the status of the switch with the value of $S(x)$:

$$\gamma = \begin{cases} 0, \text{ for } S(x) > 0 \\ 1, \text{ for } S(x) < 0. \end{cases}$$  \hspace{1cm} \text{(18)}

The existence condition equation (17) can be expressed in the form

$$S(x) = K^T Ax + K^T D < 0, S(x) > 0$$  \hspace{1cm} \text{(19)}

$$S(x) = K^T Ax + K^T B + K^T D > 0, S(x) < 0.$$  \hspace{1cm} \text{(20)}

From a simulation or practical point of view, assuming that error variables $x_i$ are suitably smaller than references $V^*$, equation (15) and equation (16) can
be rewritten in the form
\[ K^T D < 0, S(x) > 0 \] (21)
\[ K^T B + K^T D > 0, S(x) < 0. \] (22)

By substituting matrices B and D in equation (20) and equation (21), one obtains
\[ \frac{K_1}{L_1} V_{C1ref} - \frac{K_2}{C_1} i_{L1ref} - \frac{K_3}{L_2} V_{C2ref} + \frac{K_4}{RC_2} [i_{L2ref} R - V_{C2ref}] < 0 \] (23)
\[ \frac{K_1}{L_4} V_m - \frac{K_2}{C_4} i_L + \frac{K_3}{L_2} [V_m - V_{C1} - V_{C2ref}] + \frac{K_4}{RC_2} [i_{L2ref} R - V_{C2ref}] > 0 \] (24)

The existence condition is satisfied if the inequalities equation (21) and equation (22) are true.

Finally, it is necessary to guarantee that the designed sliding plane is reached for all initial states. If the sliding mode exists, in the system defined by equation (6), it is a sufficient condition that coefficients $K_1, K_2, K_3$ and $K_4$ be nonnegative.

### 3.2 Switching Frequency

![Fig. 5. Switching function $\gamma$.](image)

In the ideal sliding mode at infinite switching frequency, state trajectories are directed toward the sliding surface and move exactly along it. A practical system cannot switch at infinite frequency. Therefore, a typical control circuit features a practical relay, as indicated in Fig. 5. A practical relay always exhibits hysteresis modeled by

\[ \gamma(s) = \begin{cases} 0, & \text{when } S > +\delta \text{ or } S < -\delta \text{ or } |S| < \delta \\ 1, & \text{when } S < -\delta \text{ or } \text{when } S > +\delta \end{cases} \] (25)

Where $\delta$ is an arbitrarily small positive quantity and $2\delta$ is the amount of hysteresis in $S(x)$. The hysteresis characteristic makes it impossible to switch the control on the surface $S(x) = 0$. As a consequence, switching occurs on the lines $S = \pm \delta$, with a frequency depending on the slopes of $i_{L1}$ and $i_{L2}$. This hysteresis causes phase plane trajectory oscillations of width $2\delta$, around the surface $S(x) = 0$ as shown in Fig. 6. Note that Fig. 5 simply tells us that in $\Delta t_1$ function $S(x)$ must increase from $(-\delta$ to $\delta$) ($S > 0$), while in $\Delta t_2$ function $S(x)$ must decrease from $+\delta$ to $-\delta$ ($S < 0$).

![Fig. 6. The waveform of $S(x)$.](image)

The switching frequency equation is obtained from Fig. 6, by considering that the state trajectory is invariable, near to the sliding surface $S(x) = 0$ and is given by

\[ f_s = \frac{1}{\Delta t_1 + \Delta t_2} \] (26)

Where $\Delta t_1$ is conduction time of the switch $S$ and $\Delta t_2$ is the off time of the switch $S$. The conduction time $\Delta t_1$ is derived from equation (24) and it is given by

\[ \Delta t_1 = \frac{2\delta}{\frac{K_1}{L_1} V_m - \frac{K_2}{C_1} i_L + \frac{K_3}{L_2} [V_m - V_{C1} - V_{C2ref}] + \frac{K_4}{RC_2} [i_{L2ref} R - V_{C2ref}]} \] (27)

The off time $\Delta t_2$ is derived from equation (23), and it is given by
\[ \Delta t_2 = \frac{-2\delta}{K_2 C_1 i_{L1ref} - K_1 i_{L1ref} - K_2 V_{C1ref} + \frac{K_1}{L_2} V_{L2ref}} \]  

(28)

The maximum switching frequency is obtained substituting equation (27) and equation (28) in equation (26) in the assumption that the converter is operating with non-load (\(i_{L1ref} = 0\), \(i_{L2ref} = 0\) and \(I/R = 0\)) and the output voltage reference is passing by maximum \(V_{C1ref \, \text{(max)}}\) and \(V_{C2ref \, \text{(max)}}\). The maximum switching frequency is obtained as equation (29)

\[ f_s(\text{max}) = \frac{K_1 K_2 V_{in}}{2\delta L_1 L_2 \left(1 - \frac{V_m}{V_{C1ref \, \text{(max)}} + V_{C2ref \, \text{(max)}}}\right)} \]  

(29)

### 3.3 Duty Cycle

The duty cycle \(d(t)\) is defined by the ratio between the conduction time of the switch \(S\) and the switch period time, as represented by

\[ d(t) = \frac{\Delta t_1}{\Delta t_1 + \Delta t_2} \]  

(30)

Considering the sliding mode control an instantaneous control, the ratio between the output and the input voltages must satisfy in any working condition.

\[ \frac{V_o}{V_{in}} = \frac{1}{1 - d(t)} \]  

(31)

### 3.4 Inductor Currents

The high-frequency or maximum inductor current ripple is obtained from Fig. 2 and given by [15]

\[ \Delta i_{L1} = \frac{\Delta t_1 V_{in}}{L_1} \]  

(32)

\[ \Delta i_{L2} = \frac{\Delta t_1 V_{in}}{L_2} \]  

(33)

### 3.5 Capacitor Voltages

The controller operates over the switch to make the capacitors voltage \(V_{C1}(t)\) and \(V_{C2}(t)\) follow its reference. Over \(V_{C1}(t)\) and \(V_{C2}(t)\), a high-frequency ripple (switching) is imposed, which is given by [15]

\[ \Delta V_{C1}(t) = \frac{V_{C1}}{RC_1} \Delta t_1 \]  

(34)

\[ \Delta V_{C2}(t) = \frac{V_{C2}}{RC_2} \Delta t_1. \]  

(35)

It is interesting to note that the switching frequency, inductor current ripple, and capacitor voltage ripple depend on the following terms: the control parameters, circuit parameters, reference voltage, output capacitor voltage \(V_{C2}(t)\), pump capacitor voltage \(V_{C1}(t)\), and inductor currents \(i_{L1}, i_{L2}\).

It is important to determine the circuit parameters and coefficients \(K_1, K_2, K_3\) and \(K_4\) that agree with desirable values of maximum inductor current ripple, maximum capacitor voltage ripple, maximum switching frequency, stability, and fast response for any operating condition.

### 4. Design Computation of Circuit Components and Controller Parameters

The main purpose of this section is to use the previously deduced equations to calculate the POELC components value and controllers’ parameters.

#### 4.1 Calculation of \(V_{C2}\)

From equation (28) and a simulation point of view, the output voltage is chosen to produce a duty cycle close to 0.75. The adopted value of the output voltage \(V_o\) is 36 which is in Table 1, and a variation of the duty cycle between \(d_{\text{min}} = 0.4\) and \(d_{\text{max}} = 0.9\) is expected. Finally \(V_{C2\, \text{max}} = 108\) V.

#### 4.2 Determination of Ratio \(K_1 / L_1\) and \(K_3 / L_2\)

Substituting \(V_{in}, V_{C2\, \text{ref}} = V_{C2\, \text{max}},\) and \(\delta = 0.75\) in equation (29), the value of \(K_1 / L_1\) and \(K_3 / L_2\) is obtained as \(K_1 / L_1 = 15000\). There are some degrees of freedom in choosing the ratio \(K_1 / L_1\) and \(K_3 / L_2\). It is recommendable to choose the ratio \(K_1 / L_1\) and \(K_3 / L_2\) to agree with required levels of stability and response speed. The final adopted value is, \(K_1 / L_1\) and \(K_3 / L_2 = 1200\).

#### 4.3 Determination of Ratio \(K_2 / C_1\) and \(K_4 / C_2\)

From equation (23) and equation (24) and taking \(i_{L1ref} = i_{L1\, \text{max}} = 2.16\) A and \(i_{L2ref} = i_{L2\, \text{max}} = 2.3\) A, one obtains \(12008 < K_2 / C_1 < 2848433\) and \(12008 < K_4\).
There are some degrees of freedom in choosing the ratio \( K_2 / C_1 \) and \( K_4 / C_2 \). In this controller, the ratio \( K_2 / C_1 \) and \( K_4 / C_2 \) is a tuning parameter. It is recommendable to choose the ratio \( K_2 / C_1 \) to agree with required levels of stability and response speed. The ratio \( K_2 / C_1 \) and \( K_4 / C_2 \) is chosen by iterative procedure (i.e. the ratio is modified until the transient response is satisfactory), and it is verified by simulation. The final adopted value is, \( K_2 / C_1 = 40000 \).

### 4.4 Calculation of \( L_1 \) and \( L_2 \)

The maximum inductor current ripple is chosen to be equal to 15% of maximum inductors current and the inductors value which is obtained from equation (32) and Eq. (33) as \( L_1 = 1 \text{mH} \).

### 4.5 Calculation of \( C_1 \) and \( C_2 \)

The maximum capacitor ripple voltage \( \Delta V_{c1\text{max}} \) and \( \Delta V_{c2\text{max}} \) is chosen to be equal to 1% maximum capacitors voltage and capacitors \( C_1 \) and \( C_2 \) are determined using equation (34) and equation (35) as \( 20 \mu\text{F} \).

### 4.6 Values of the Coefficients \( K_1, K_2, K_3 \) and \( K_4 \)

Having decided on the values of the ratio \( K_1 / L_1 \) and \( K_3 / L_2 \) inductors, the value of \( K_1 \) and \( K_3 \) is unswervingly obtained (\( K_1 = 1.2 \) and \( K_3 = 1.2 \)). Similarly the \( K_2 \) and \( K_4 \) (\( K_2 = 0.8 \) and \( K_4 = 0.8 \)) is computed using the ratio \( K_2 / C_1 \) and \( K_4 / C_2 \) and capacitors \( C_1 \) and \( C_2 \).

### 5. Simulation Study

The main purpose of this section to discuss about the simulation studies of the POELC with designed SMC. The validation of the system performance is done for five regions viz. transient region, line variations, load variations, steady state region and also components variations. The simulations have been performed on the POELC circuit with parameters listed in Table 1.

The static and dynamic performances of SMC for POELC are evaluated in MatLab/Simulink software platform. The detailed operation of SMC for POELC is presented in sections 2 and 3.

**Table 1**

Parameters of the POELC.

<table>
<thead>
<tr>
<th>Parameters name</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage</td>
<td>( V_{in} )</td>
<td>12V</td>
</tr>
<tr>
<td>Output Voltage</td>
<td>( V_{o} )</td>
<td>36V</td>
</tr>
<tr>
<td>Inductors</td>
<td>( L_1, L_2 )</td>
<td>1mH</td>
</tr>
<tr>
<td>Capacitors</td>
<td>( C_1, C_2 )</td>
<td>20 ( \mu )F</td>
</tr>
<tr>
<td>Nominal switching frequency</td>
<td>( f_s )</td>
<td>100kHz</td>
</tr>
<tr>
<td>Load resistance</td>
<td>( R )</td>
<td>50 ( \Omega )</td>
</tr>
<tr>
<td>Output power</td>
<td>( P_o )</td>
<td>25.92W</td>
</tr>
<tr>
<td>Input power</td>
<td>( P_{in} )</td>
<td>25.92W</td>
</tr>
<tr>
<td>Input current</td>
<td>( I_{in} )</td>
<td>2.16A</td>
</tr>
<tr>
<td>Range of duty ratio</td>
<td>( d )</td>
<td>0.4 to 0.9</td>
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<tr>
<td>Adopted value of duty ratio</td>
<td>( d )</td>
<td>0.75</td>
</tr>
</tbody>
</table>
5.1 Transient Region

Fig. 7 shows the dynamic behavior in the startup for output voltage of POELC for different input voltages like 5V, 12V and 19V respectively. It can be seen that output voltage of POELC for Vin = 5 V, 12 V, and 19 V has negligible overshoot and settling times of 0.1 s, 0.04 s and 0.02 s in startup with designed SMC.

Fig. 8 shows the dynamic behavior in the startup for output voltage of POELC for various load resistances like 10 Ω, 50 Ω, and 90 Ω respectively. It can be seen that output voltage of POELC has slightly overshoot and settling time of 0.028 s for R = 100 Ω, whereas the output voltage of POELC for R = 50 Ω and 90 Ω has negligible overshoot and settling times of 0.03 s and 0.17 s in startup with designed SMC.

5.2 Line Variations

Fig. 10 shows the response of output voltage of POELC for input step change from 12V to 15 V (+30% line variations). It can be found that the output voltage of POELC has maximum overshoot of 4

Fig. 9 represents a simulation of the dynamic behavior in the state plane for the input current and output voltage of POELC with designed SMC in sliding surfaces. It can be found that input current of POELC goes up to 2.16 A and output voltage of POELC travels up to 36 V without overshoot.
V and settling time of 0.038 s with designed SMC.

Fig. 11. Response of output voltage of POELC for input step change from 12V to 9V.

Fig. 11 shows the response of average output voltage of POELC for input voltage step change from 12 V to 9 V (-30% line variations). It can be found that the output voltage of POELC has maximum overshoot of 4 V and settling time of 0.038 s with designed SMC.

5.3 Load Variations

Fig. 12. Response of output voltage of POELC when load value takes a step changes from 50 Ω to 60 Ω.

Fig. 12 shows the response of output voltage of POELC for load step change 50 Ω to 60 Ω (+20% load variations). It could be seen that the output voltage of POELC has maximum overshoot of 3.8 V and settling time of 0.038 s with proposed control scheme.

5.4 Steady State Region

Fig. 13. Response of output voltage of POELC when load value takes a step changes from 50 Ω to 40Ω.

Fig. 13 shows the response of output voltage of POELC for load step change 50 Ω to 40 Ω (-20% load variations). It could be seen that the output voltage of POELC has maximum overshoot of 3.8 V and settling time of 0.038 s with proposed control scheme.

Fig. 14. Response of output voltage and inductor current \(i_{L1}\) in steady state condition.

Fig. 14 shows the instantaneous output voltage and inductor current \(i_{L1}\) of POELC with proposed control scheme in steady state region. It is evident from the figure that the output voltage ripple is very small 0.03V and peak-to-peak inductor current ripple is 0.18 A for the average switching frequency (100 kHz) closer to theoretical designed value listed in Table 1.
5.5 Components Variation

Fig. 15. Output voltage when inductors ($L_1$ & $L_2$) variation from 1mH to 2mH.

Fig. 16. Output voltage when capacitors ($C_1$ & $C_2$) variation from 20 µF to 50 µF.

Fig. 17. Average input and output current of POELC.

Fig. 18. Simulated measured average output voltage of POELC at $V_{in}$=12 V and load $R$ = 10 Ω, 40 Ω and 50 Ω for different hysteresis band settings.

Fig. 15 shows the output voltage of POELC with proposed control scheme for inductor variation from 1mH to 2mH. It could be found that the change does not influence the converter behaviour due to proficient design of SMC.

An interesting result is illustrated in Fig. 16. It shows the output voltage response of POELC with proposed control scheme for the variation in capacitor values 20µF to 50µF. It can be seen that the SMC is very successful in suppressing effect of capacitance variation except that a negligible output voltage ripples.

Fig.17 shows the average input current and output current of POELC with proposed control scheme respectively. It is showed that the average input current is 2.16A and average output current is 0.72A, which is very closer to theoretical value in Table 1.

Using simulation analysis the POELC with hysteresis rule based SMC computes that the input and output power values are 25.92W and 25.92W respectively, which is very closer to the calculated theoretical value listed in Table 1.
Fig. 19. Simulated measured average switching frequency of POELC at Vin = 12 V and load R= 10 Ω, 40 Ω and 50 Ω for different hysteresis band settings.

Fig. 20. Simulated measured average output voltage of POELC for different input voltage.

Figs. 18 and 19 shows the graphs of simulated measured average output voltage and average switching frequency of POELC against hysteresis band for load resistances R = 10 Ω, 40 Ω, and 50 Ω respectively. From the figure, the average output voltage is lower with lower load resistance, and average switching frequency is notably higher with lower load resistance with designed SMC.

The simulated measured average output voltage of POELC for input voltage range 6V to 19V is plotted in Fig. 20. It can be found that the average output voltage increase with increasing input voltage with designed SMC. Specifically, output voltage deviation of POELC is 0.1 V for the entire input voltage range.

A Proportional-Integral (PI) controller with settings $K_p = 0.0112$ and $T_i = 0.0113s$ obtained by the Ziegler-Nichols tuning technique [20] has been used for comparison with the designed SMC. Figs. 21 and 22 show the simulated time domain performances evaluation of current/voltage profiles of POELC using SMC versus PI controller. From this figures it is clearly identified that the simulated results of the designed SMC are showed better performance over the PI controller under line and load disturbances in steady state operating region.

Fig. 21. Time domain performances evaluation of POELC using SMC versus PI controller under line disturbances in steady state operating region.

Fig. 22. Time domain performances evaluation of POELC using SMC versus PI controller under load disturbances in steady state operating region.
6. Conclusions

Switched mode DC-DC power converters are used in a variety of electric power supply systems, including cars, ships, aircrafts, high voltage projects and computers. Power electronic converters are intrinsically periodic time-variant structure systems due to their inherent switching operation, so the sliding mode control approach is a strong candidate method for the converter controller design.

The design and output voltage regulation of SMC for POELC operated in CCM has been successfully demonstrated in this paper. A SMC over the output capacitor voltage and inductor current has been used for the control. The influence of the control parameters on the performances of the system was studied. The effect of proper selected controller parameters of sliding mode controlled POELC operated in CCM resulted in fast dynamic response and excellent static and transient responses over the conventional controller. It is, therefore, feasible for common DC-DC conversion purpose, computer power supplies and medical equipments etc. Further research may focus to the study the experimental set-up of the proposed system.

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Biography

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